

Amendments to the Specification:

Please amend paragraph [0002] as shown below:

[0002] The present invention relates to an extended resonance based phased array system for reducing and/or eliminating the need of a separate power splitter and phase shifter shifters in a conventional phased array system, which results in a very compact and simple circuit structure at lower-cost.

Please delete paragraph [0004] as shown below:

~~[0004] Recently, several new beam steering techniques have been demonstrated, which attempt to reduce the known problems with phase arrays. The techniques demonstrated rely on piezoelectrically actuated mechanical systems to achieve phase shifting. In another demonstration, the dielectric tunability of a ferroelectric based lens is used to achieve beam steering. In yet another demonstration, by changing the frequency of an injection signal to an array of injection-locked oscillators, beam steering is achieved.~~

Please amend paragraph [0007] as shown below:

[0007] A phased array is a group of antennas in which the relative phases of the respective signals feeding the antennas are varied electronically in such a way that the effective radiation pattern of the array is reinforced in a desired direction and suppressed in undesired directions. Phased arrays are the ideal solution for many applications, such as early warning and missile defense systems, satellite communications, traffic control systems, automotive collision avoidance and cruise control systems, blind spot indicators, compact scanning arrays, smart base station antennas for cellular communications, etc. In a conventional phased array, the signal is divided into many branches using a corporate feed network and each branch is then fed into a phase shifter and followed by an antenna. Phase shifters are considered as

the most sensitive and expensive part of a phased array. Also, the complexities in the corporate feed network, the bias network for the phase shifters, and the interactions between array elements, can make the design of phased arrays very challenging and expensive. Therefore, the phased arrays have been used only in a few sophisticated military applications and space systems. These applications usually have stringent requirements on the side lobe levels, scan range and beam width of the phased arrays.

On the other hand, phased arrays are being considered for emerging commercial applications, such as automotive collision avoidance systems, mobile multimedia broadcasting, and traffic control radars. In these systems, accurate beam control and wide scan angle are not required. Instead, low cost, small size, and ease of manufacturability are the driving criteria.

Please amend paragraphs [0009] – [0013] as shown below:

[0009] The present invention can provide dramatic cost reductions in the cost of phased array antenna systems. As discussed earlier, phased arrays based on this technique do not need separate power splitter and phase shifters. The phased arrays according to the present invention simply use varactors (i.e. capacitors devices whose capacitance can be varied with an applied DC voltage) for splitting the power and achieving the required phase shift. ~~A price comparison can be made between the cost of phase shifters in a conventional phased array and the cost of tunable capacitors required to design the phased arrays based on the technique according to the present invention.~~

~~[0010] Phase shifters are typically constructed using ferrite materials, p-I-n diodes, or field effect transistor (FET) switches. Ferrite-based phase shifters exhibit low loss, but the size and cost make the ferrite-based phase shifters prohibitively expensive for phased array applications. Solid-state (pin diode or FET) based phase shifters are extensively used in modern phased array systems, but the solid-state based phase shifters suffer from significant amount of loss and require additional amplification to compensate for the loss, which increases the cost. Nowadays, research activities concentrate on micro-electro-mechanical systems (MEMS) and~~

~~ferroelectric-based phase shifters to address these issues. The table below shows the approximate prices of commercially available solid-state based phase shifters:~~

Frequency Band	Approximate price range per phase shifter
L-band (1-2 GHz)	US \$57*
X-band (8-12 GHz)	US \$46—US \$102*
Ku-band (12-18 GHz)	US \$46—US \$102*
Ka-band (27-40 GHz)	US \$70—US \$145*

~~*Prices shown were taken from commercial phase shifter suppliers including MACOM, Triquint Semiconductor, TLC Precision Wafer Technology and KDI Corporation for a quantity of 1000 phase shifters.~~

[0011] As mentioned earlier, phased arrays based on the technique of the present invention use tunable capacitors, or varactors. Varactors can be fabricated based on solid-state, MEMS, and ferroelectric technologies. The solid-state based varactors are well-mature and can easily be obtained commercially, whereas the MEMS and ferroelectric based varactors are still under development. ~~Varactors can cost anywhere between US \$1 and US \$10 depending on the capacitance of the varactor, tuning range and quality factor.~~

[0012] ~~For comparison, a linear phased array of ten antennas can be considered. The linear phased array of ten antennas needs ten phase shifters, if built using the conventional approach and the phase shifters cost approximately US \$800 (i.e. the cost of a phase shifter is assumed to be US \$80), whereas the linear phased array of ten antennas needs 20 varactors, if built using the technique according to the present invention and the varactors cost approximately US\$100 (i.e. the cost of a varactor is assumed to be US \$5). This implies more than a 50% reduction in cost compared to the cost of phase shifters in a conventional system. The reduction in the cost becomes even more significant as the order of the array increases.~~

[0013] Phased arrays have been finding increasing number of applications in military and commercial communication systems. The phased array system can steer

a beam rapidly by electronically tuning the relative phase between the antennas compared to mechanical beam-steering. Mostly, ~~ferrite or semiconductor based phase shifters are employed to tune the phase difference between antennas.~~ Conventional phased array use a phase shifter for each antenna element. However, the cost of the phased array increases significantly with the number of phase shifters used. These systems are also very complex, bulky and heavy and ~~suffer from high loss and mass.~~ Cost reduction and performance improvement is necessary in phased arrays to address ~~follow~~ the emerging commercial applications, such as smart antennas, automotive collision avoidance and cruise control systems.

Please amend paragraphs [0015] – [0016] as shown below:

[0015] Tunable capacitors can be based on varactor diodes, ferroelectric tunable capacitors, MEMS tunable capacitors or adjustable length of transmission lines using various switches like PIN diodes, transistors, ~~mechanical or MEMES~~ MEMS switches.

[0016] Tunable inductors can be based on ferrite devices or active inductors (use using transistors to emulate inductors). Some of the applications of the PDPS circuits include: (1) Low cost one and two dimensional phased array antennas; (2) Tunable transversal active filters; and (3) Tunable transversal equalizers.

Please amend paragraph [0019] as shown below:

[0019] FIG. 1 illustrates an extended resonance concept incorporating N-ports according to one embodiment of the invention ~~based array system according to the present invention incorporating N-devices;~~

Please amend paragraphs [0042] – [0045] as shown below:

[0042] FIG. 24 illustrates a measured H-plane pattern for various diode voltages according to the present invention, where measured gain at 30 V is 8.7 dB; and

[0043] FIG. 25 illustrates a measured return loss for various diode voltages according to the present invention;

[0043.1] FIG. 26 is a detailed illustration of an embodiment where a second tunable element is a switching fixed capacitor C configuration to be inserted in place of the second tunable element illustrated in any of FIGS 1-3 or FIGS 13-15; and

[0043.2] FIG. 27 is a detailed illustration of an embodiment where a second tunable element is a switching transmission line l configuration to be inserted in place of the second tunable element illustrated in any of FIGS 1-3 or FIGS 13-15.

DESCRIPTION OF THE PREFERRED EMOIDIMENT

[0044] The present invention uses extended resonance which is a power dividing/combining technique, which has been exploited for the design of power amplifiers at microwave and millimeter wave frequencies. It results in very compact structures with high dividing/combining efficiency (>90%) up to millimeter wave frequencies. An N-port extended resonance dividing circuit is shown in FIG. 1. The admittance of the first and the last port is $G+jB$ (where G is conductance and B is susceptance), whereas the admittance of the each interior port is $G+2jB$. The length of the transmission line, l_1 , is chosen such that the admittance of the first port is transformed to its conjugate, $G-jB$. The admittance at the plane of the second port will be $2G+jB$. As can be seen, half of the susceptance of the second device is cancelled in this process. The length of the second transmission line, l_2 (not shown in FIG. 1), is chosen to transform $2G+jB$ to its conjugate, $2G-jB$ (not shown in FIG. 1). The admittance at the plane of the third port will be $3G+jB$ (not shown in FIG. 1). This process is performed (N-1) times. At the last stage, the admittance at the plane of the (N-1)th transmission line, l_{N-1} , will be (N-1) $G-jB$ and the admittance at the plane of the Nth port will be NG , which is matched to the source impedance R_s using a quarter-wave transformer, $\lambda/4$. Resonating all the ports with one another essentially places the ports in shunt, and analysis of this structure shows that the voltage at each port is equal in magnitude, but generally not in phase. This feature

has been exploited for the design of power amplifiers at microwave and millimeter wave frequencies. It can be shown that by correct selection of susceptance B and conductance G , one can maintain equal power division, and vary the relative phase shift between device nodes by changing susceptance B . It should also be mentioned that it is possible to design an extended resonance dividing circuit for arbitrary imaginary part of the port admittances as long as the admittances are transformed to their conjugates and the last stage is matched to the source impedance R_s .

[0045] The concept of a phased array based on the extended resonance technique can be explained as follows: The port in FIG. 1 compared to FIG. 2 is modeled as a shunt combination of an antenna ($G=G_{ant}$) and a capacitor ($B=j\omega C$). An inductor L , in FIG. 2, is used to transform the admittance to its conjugate instead of a transmission line l , as used in FIG. 1. A schematic illustration of the proposed phased array is shown in FIG. 2. The antennas are assumed to be $\lambda/2$ apart, and the capacitors C and inductors L are assumed to be tunable. It can be shown that the required inductance to transform the admittance, $nG_{ant}+j\omega C$, to its conjugate, $nG_{ant}-j\omega C$, is:

$$L_n = \frac{2C}{(nG_{ant})^2 + (\omega C)^2} \quad (1)$$

Please amend paragraphs [0048] - [0049] as shown below:

[0048] It can be concluded from equation (3) that changing the capacitance at each port will result in a change in the phase difference between the successive antenna ports. In a phased array, the phase shifts between successive antenna ports must be equal to each other ($\theta_{21} = \theta_{32} = \theta_{43} \dots$). Depending on the number of antennas, N , and the tunability of the capacitor, there exists an optimum capacitive susceptance, which results in the same phase shift between the successive antenna nodes while dividing the power equally. Therefore, a phased array system with one dimensional scanning capability can be built. Since realizing tunable inductors is not very easy and the antennas have to be spaced approximately $\lambda/2$ apart depending on

the design, the circuit of FIG. 2 may not be practical. Instead, artificial tunable inductors can be realized using an impedance inverter consisting of two quarter-wave transformers $z_0, \lambda/4$ with a shunt tunable capacitor C_L in between. Phase offsets must be introduced prior to the antennas to make the absolute phases of the voltages at the antenna ports equal to each other. The proposed extended resonance based phased array system is shown in FIG. 3. Based on the theory outlined, the simulated normalized radiation pattern for a five antenna extended resonance based phased array at 2 GHz for various capacitor values, C , is shown in FIG. 4. In this simulation, no loss from the tunable capacitors or transmission lines is included. The simulated maximum scanning range for various array sizes as a function of the capacitor tunability is plotted in FIG. 5. It can be concluded that for this particular design, the maximum achievable scan range is approximately 44 degrees. The effect of the capacitor quality factor on the array efficiency is also shown in FIG. 6. It turns out that with a moderate capacitor quality factor ($Q \sim 10$), it is possible to obtain higher than 80% efficiency. Extended resonance based phased arrays can reduce and/or eliminate the need for a separate power splitter and phase shifters in a conventional phased array system, which results in a compact, simple and low-cost circuit architecture.

EXAMPLE 1

[0049] To demonstrate the operation of this technique, a two GHz extended resonance based phased array including four edge coupled microstrip patch antennas placed half wavelength apart was designed, fabricated and tested. A 31 mil thick RT/dureid RT/DUROIDTM 5880 substrate from Rogers Corporation was used to build the phased array. MSV34 series chip varactor diodes from Metelics Inc. were used as tunable capacitors. A photo of the phased array can be seen in FIG. 7. The overall size of the phased array was $39 \times 25 \text{ cm}^2$. The measured H-plane pattern of the phased array for various diode voltages is shown in FIG. 8 and the measured performance is summarized in Table 1. The results show that the phased array can scan the beam ± 13.5 degrees with the application of 2 V to 30 V reverse bias to the varactor diodes. The side lobe level was better than 7 dB. The gain of the phased array was measured to be 8.3 dB at 30 V reverse bias applied to the varactors. It can

be seen from FIG. 8 that the gain at 2 V is 6.9 dB lower than the gain at 30 V. This is due to the low quality factor of the varactor diodes at this voltage ($Q_{2V}=22$, $Q_{30V}=121$ at 2 GHz), resulting in significant amount of RF power dissipation within the diode and change in the input impedance, which degrades the return loss. It should be noted that any type of tunable capacitors, such as ferroelectric or MEMS based tunable capacitors, switched capacitors using PIN diodes or MEMS switches, which have been known to have lower loss, can be used to fabricate the phased array. In extended resonance based phased arrays, fewer number of devices are employed compared to a conventional phased array system, thereby reducing the cost.

Please amend paragraphs [0051] – [0054] as shown below:

[0051] The concept of extended resonance based phased arrays is shown in FIG. 12. The concept uses tunable capacitors \underline{C} and tunable inductors \underline{L} . The admittance seen at the plane of the 1st port ($G_{ant}+j\omega C$) is transformed to its conjugate ($G_{ant}-j\omega C$) using the 1st inductor (\underline{L}_1). Similarly, the admittance at the 2nd port ($2G_{ant}+2j\omega C$) is transformed to its conjugate using the 2nd inductor (\underline{L}_2). This process is performed (N-1) times, and the admittance seen at the plane of the last port will be NG_{ant} , which is matched to the source impedance using a matching network. The analysis of this structure shows that the voltages at each port are equal in magnitude (equal power division among antennas), and the phase difference between adjacent ports are all equal to each other. Therefore, by tuning the varactors as well as inductors, one can obtain equal power division among antennas and phase shifting between successive ports. Thus, a phased array system with one-dimensional scanning capability can be designed. Due to the initial phase offsets between the power divider ports, constant phase delays ($\Phi_{offset1}$, $\Phi_{offset2}$, . . . $\Phi_{offsetN}$) are used as shown in FIG. 12 to set the initial phases at the antenna nodes equal to each other. From then on, the beam is steered around the boreside of the antennas by tuning the varactors. It should also be noted that an extended resonance circuit can be designed for a specified amplitude taper to achieve low side lobe. Since the magnitude of the voltage \underline{V} is always the same as long as the admittances seen at the ports are

transformed to their conjugates, non-uniform amplitude distribution can be obtained by adjusting the conductances seen at the ports (or antenna input impedances). In some designs unequal power distribution is desirable, for example arrays using Chebyshev tapered distribution for lower side lobes. The design according to the present invention can accommodate this.

[0052] Tunable inductors were previously realized using impedance inverters consisting of two quarter-wave transformers $z_0, \lambda/4$ with a shunt varactor C_1 in between, as shown in FIG. 3. However, this approach has a bandwidth limitation due to the quarter-wave transformers used. Furthermore, the structure of FIG. 4 requires the value of the tunable capacitors C to increase progressively as odd multiple of the first varactor capacitance and the value of the tunable inductors L to decrease progressively compared to the first inductor. This can place a limit on the design of varactors and possible capacitance values available. In this section, the design methodology of an extended resonance based phased array, which uses fixed inductors and single value varactors is presented.

[0053] The required inductance to transform the admittance, $nG_{ant} + nj\omega C$, to its complex conjugate, $nG_{ant} - nj\omega C$, is [6]:

$$L_n = \frac{2C}{nG_{ant}^2 + n\omega^2 C^2} \quad (4)$$

[0054] Using equation 4 (and assuming $\omega C_{max} = G_{ant} \sqrt{t}$ for maximum phase shift), the required tunability for the tunable inductors is calculated as:

$$t_L = \frac{1+t}{2\sqrt{t}} \quad (5)$$

where t is the tunability of the varactor (the ratio of the maximum capacitance to the minimum capacitance, $t = C_{max}/C_{min}$). The required tunability for the inductors increases as the tunability of the varactors increase, but not at the same rate. For example, $t_L = 1.34$ for a varactor with $t = 5$ and $t_L = 1.74$ for a varactor with $t = 10$. Since

not much tunability is required for the inductors, in this design, the value of the inductor is kept constant at an average value between its maximum and minimum values at the expense of tolerating some small power division and phase errors (see FIG. 4.). Consider a generalized extended resonance phased array circuit in FIG. 2. $P_1, P_2, \dots P_N$ designate the required powers going into the antennas to achieve a specified amplitude taper. Since the magnitude of the voltage between power divider ports are equal to each other, the conductances seen at the power divider ports (or input conductances of the antennas) are designed to achieve the required power ratios. For example, the 2nd conductance will be

$$G_2 = G_1 P_2 / P_1 \quad (6)$$

Please amend paragraphs [0056] - [0057] as shown below:

[0056] Similarly, this process is performed N-1 times, and at the last stage, the real admittance is matched to the source impedance using a matching network. Since amplitude coefficients for a phased array are usually symmetric, the structure of FIG. 2 is further modified as shown in FIG. 3. Half of the phased array can be designed for the desired amplitude coefficients, and two of these phased arrays are connected using an extended resonance network. This structure will have several advantages over the structure in FIG. 2, such as reduced frequency scanning due to its symmetry, and physically realizable matching networks. The left and right ~~part~~ portions, with respect to a common combined input line corresponding to a line of symmetry joining the two phased arrays (not shown), of the phased arrays must be isolated while biasing the varactors. The varactors on the left side portion and the right side portion must be biased such that the same progressive phase shift is obtained between successive ports compared to the phase shift when the phased array scans the boreside. Based on the theory outlined, simulated array factor for an X-band 8-antenna phased array is shown in FIG. 4. In this simulation, the varactor quality factor is assumed to be 15 at 10 GHz, and inductors are kept constant. The phased array can steer the beam 31 degrees by tuning the varactors between 0.9 pF

and 0.159 pF. The side lobe levels are better than 15 dB. This degradation compared to the designed side lobe level of 20 dB is due to utilization of the fixed inductors.

EXAMPLE 2

[0057] A 10 GHz extended resonance based phased array including 8 microstrip patch antennas has been designed, fabricated and tested. The antennas were half wavelength apart. A 15 mil thick TMM3TM substrate from Rogers Corporation was used to build the phased array. MA46580 series beam lead varactor diodes from MACOM Inc. were used as tunable capacitors. A photo of the phased array is shown in FIG. 9. The overall size of the phased array was $11.4 \times 3 \text{ cm}^2$ (except for the bias lines and input feed line). The measured H-plane pattern of the phased array for various diode voltages is shown in FIG. 10. The preliminary measurement results show that the phased array can steer the beam 18 degrees with the application of 2.25 V to 10.2 V reverse bias to the varactor diodes. The measured side lobe level was better than 10 dB. It can be seen from FIG. 10 that the gain of the phased array decreases as the diode voltage is reduced to 2.25 V. This is due to the low quality factor of the varactor diodes at this voltage, resulting in significant amount of RF power dissipation within the diode and change in the input impedance, which degrades the return loss. In extended resonance based phased arrays, fewer number of devices are employed compared to a conventional phased array system, thereby reducing the cost. The circuit topology presented here also simplifies the design of large phased arrays while having a compact circuit area for dividing the power and phase shifting.

Please amend paragraphs [0060] – [0061] as shown below:

[0060] A modified approach with improved performance is disclosed in the present invention. An N-port extended resonance power divider circuit is shown in Fig. 13. The admittance connected to the n^{th} port for $n < N$ is $G + 2(n-1)jB$, whereas the admittance connected to the last port is $G + (N-1)jB$. The length of the first transmission line, l_1 , is chosen such that the admittance at the first port is transformed to its conjugate, $G - jB$. The admittance seen at the second port is $2(G + jB)$. Similarly,

the length of the second transmission line, l_2 (not shown in FIG. 13), is chosen to transform $2(G+jB)$ to its conjugate, $2(G-jB)$, hence the admittance seen at the third port is $3(G+jB)$. This process is performed $(N-1)$ times, and at the last stage, the admittance seen at the plane of the $(N-1)^{\text{th}}$ transmission line will be $(N-1)(G-jB)$ and the admittance seen at the N^{th} port will be NG , which is matched to the source impedance using a quarter-wave transformer. The analysis of this structure shows that the voltages at each port are equal in magnitude (equal power division), but not in phase. This feature has been exploited for the design of power amplifiers at microwave and millimeter wave frequencies.

[0061] The concept of a phased array based on the extended resonance technique is depicted in FIG. 14. The power divider ports are connected to an antenna ($G=G_{\text{ant}}$) in shunt with a tunable capacitor (varactor) ($B=\omega C$). Instead of a transmission line l , a tunable inductor L is used to transform the admittance to its complex conjugate as the shunt varactors are tuned. The required inductance to transform the admittance, $nG_{\text{ant}}+n\omega C$, to its complex conjugate, $nG_{\text{ant}}-n\omega C$, is:

$$L_n = \frac{2C}{nG_{\text{ant}}^2 n\omega^2 C^2} \quad (8)$$

Using the inductor value found in equation (8), the ratio of the voltages between successive ports is:

$$\frac{V_{n+1}}{V_n} = \frac{(G_{\text{ant}} + j\omega C)^2}{G_{\text{ant}}^2 + \omega^2 C^2} \quad (9)$$

Therefore, the magnitude of the voltage ratio is

$$\left| \frac{V_{n+1}}{V_n} \right| = 1 \quad (10)$$

and the phase difference between successive ports is

$$\angle \frac{V_{n+1}}{V_n} = \theta_{n+1,n} = \tan^{-1} \left\{ \frac{2\omega C G_{ant}}{G_{ant}^2 - \omega^2 C^2} \right\} \quad (11)$$

Equation (11) can be further simplified as:

$$\theta_{n+1,n} = 2 \tan^{-1} \left\{ \frac{\omega C}{G_{ant}} \right\} \quad (12)$$

Note that the phase differences between successive power divider ports given by equation (12) are all equal to each other regardless of the port number in the circuit. It should be mentioned that in a uniform amplitude phased array, the amplitude of the signal at the antennas must be the same and the phase of the signal at each antenna must successively change by the same amount. Therefore, by tuning the varactors as well as inductors given by equation (8), one can obtain equal power division among antennas as given in equation (10) and the same phase shift between successive power divider ports as given in equation (12). Thus, a phased array system with one-dimensional scanning capability can be designed. It should also be noted that an extended resonance circuit can be designed for arbitrary real and imaginary parts of the port admittances as long as the admittances seen at the ports are transformed to their conjugates. In that case, the magnitude of the voltage at each port will be equal to each other and non-uniform power distribution among antennas will be obtained to achieve low side lobe. Due to the initial phase offsets between the power divider ports, constant phase delays ($\Phi_{offset1}, \Phi_{offset2} \dots \Phi_{offsetN}$) are used as shown in FIG. 14 to set the initial phases at the antenna nodes equal to each other. From then on, the beam is steered around the boreside of the antennas by tuning the varactors. Since realizing tunable inductors is not easy, the circuit of FIG. 14 can be further modified. Artificial tunable inductors can be realized using an impedance inverter consisting of two quarter-wave transformers $\lambda/4$ with a shunt varactor C_L in between. This will both ease the realization of the tunable inductors and provide approximately $\lambda/2$ spacing for the antennas. A more realizable extended resonance based phased array

circuit is shown in FIG. 15.

Please amend paragraph [0063] as shown below:

[0063] Based on the theory outlined, simulated array factor for a 4-antenna extended resonance phased array for various normalized capacitive susceptances is shown in FIG. 18 (antennas are $\lambda/2$ apart). The simulated scan range is 21 degrees for the varactor tunability of 3.2:1. In this simulation, the varactors and transmission lines were assumed to be lossless. The effect of finite varactor quality factor (Q) on the efficiency of the extended resonance array feed has also been studied. The equivalent circuit model for the varactor is shown in FIG. 19 and its associated quality factor is given in equation (17).

$$Q = \frac{\omega C}{G_c} \quad (17)$$

where C = capacitance of a tunable capacitor, and G_c = shunt conductance of the tunable capacitor that is responsible for the loss in a nonperfect tunable capacitor. Essentially, the nonperfect tunable capacitor is modeled as a shunt combination of a lossless tunable capacitor and a shunt conductance.

Please amend paragraphs [0067] – [0068] as shown below:

[0067] To demonstrate the utility of this technique, a 2 GHz extended resonance based phased array consisting of four edge coupled microstrip patch antennas placed half wavelength apart was designed, fabricated and tested. A 31 mil thick RT/duroid RT/DUROID™ 5880 substrate from Rogers Corporation and MSV34 series chip varactor diodes from Metelics Inc. were used to fabricate the phased array. The antenna dimensions were 2.31 x 1.96 inch². The input impedance of the antenna was designed as 67 Ω by recessing the feed point by 637 mils. The tunability of the varactors was 3.2:1 with the application of 3 V to 30 V reverse bias.

Date August 21, 2008

Reply to Office Action dated February 21, 2008

A photo of the phased array is shown in FIG. 22. The overall size of the phased array is 15.4 x 9.8 inch². The radiation pattern of the phased array has been measured in an anechoic chamber, and the efficiency of its extended resonance feed was determined by measuring the magnitude and phase of the signal at each antenna node using a vector network analyzer. The measured scan angle and array feed efficiency versus the diode voltage is shown in FIG. 23. Measured H-plane patterns of the phased array for various diode voltages are also shown in FIG. 24 and the measured performance is summarized in

TABLE II
THE MEASURED PERFORMANCE OF THE PHASED ARRAY

3 dB					
Diode Voltage, V	Scan Angle, degrees	Beamwidth, degrees	Side Lobe Level, dB	Gain, dB	Efficiency, %
3	10	24	-91	6.9	59
4	6	24	-12	7.5	67
8	2	26	-14	8.1	80
10	0	24	-13.5	8.4	82
12	-2	24	-12.5	8.4	82
18	-4	26	-11	8.6	83
24	-6	26	-11	8.7	82
30	-10	28	-9	8.7	80

[0068] The phased array can steer the beam by +/-10 degrees with the application of 3 V to 30 V reverse bias to the varactor diodes, which compares well with the simulated scan range. The measured side lobe level was better than -9 dB and the average 3-dB beam width was 25 degrees. The measured array feed efficiency is typically 80% (corresponds to 1 dB insertion loss). It drops to 59% (2.3 dB insertion loss) as the diode voltage is reduced to 3 V due to the increased loss of the varactors at low reverse bias voltages. It should be noted that other tunable capacitors with lower loss, such as ferroelectric or MEMS based tunable capacitors,

switched capacitors or transmission lines using PIN diodes or MEMS switches can be utilized to fabricate the extended resonance phased arrays with better performance. The measured return loss of the phased array was better than 10 dB for all the diode voltages tested as shown in FIG. 25 and cross-polarization was lower than -23 dB.